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Underwater acoustic measurements often require wide-band sensing devices having very low self-noise levels, especially at ultrasonic frequencies. The field-effect transistor (FET) preamplifier described in this report makes possible an approximately 10-dB reduction in self noise, as compared with the best available preamplifier using other amplifying devices.

Design and development of the low-noise field-effect transistor preamplifier were performed at the Naval Undersea Warfare Center (formerly Naval Ordnance Test Station) under Bureau of Naval Weapons Task Assignment RUTO-3E-000/2161/F108-03-08 (Problem No. 115). Preliminary measurements on field-effect devices were begun in June 1965; evaluation of prototype preamplifiers was begun in December 1965. This report describes preamplifier circuit design and performance and was reviewed for technical accuracy by Guy J. Andrews and A. G. DiLoreto, Ph.D., of this Center.

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PROBLEM

Design a low-noise, field-effect transistor (FET) preamplifier, and give criteria for the low-noise operation of relevant FET devices.

RESULTS

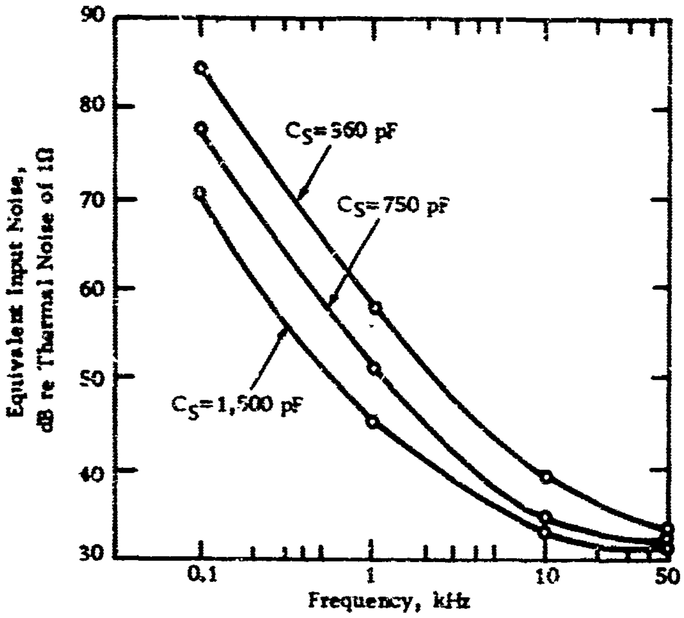
Used in conjunction with a hydrophone having a sensitivity of −80 dBV/μbar and a capacitance of 60 pF, the preamplifier described here provides a measuring system having a self-noise threshold approximately 10 dB below Sea State 0 from less than 100 Hz to nearly the thermal limit. At the same time, DC bias stability is maintained between 0°C and 50°C, and voltage gain is 6 dB with frequency response essentially flat from 1 Hz to 1 MHz. Input impedance is approximately 1,000 MΩ at low frequencies and input capacitance at ultrasonic frequencies is approximately 5 pF.

RECOMMENDATIONS

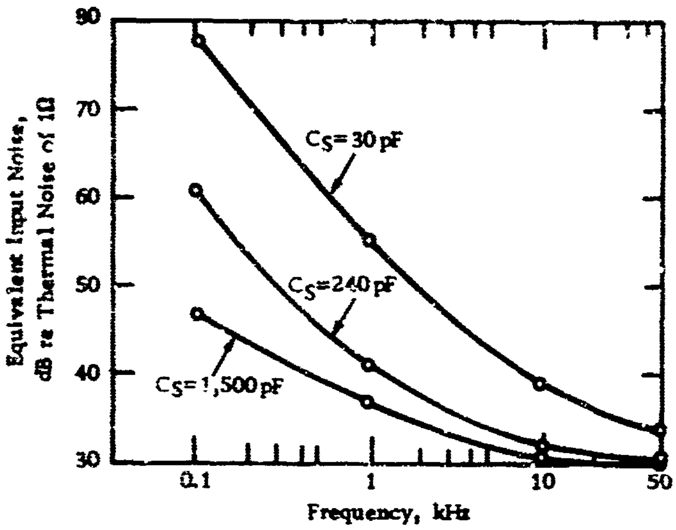
Since the preamplifier noise is primarily dependent on the noise properties of the FET, the performance of this circuit could be improved by substituting improved FET devices as they become available.

INTRODUCTION

The limitations formerly inherent in available preamplifiers have forced serious compromises in hydrophone performance, especially when sensitivity, large bandwidth, omnidirectionality, and low noise were desired simultaneously. In such cases, the preamplifier has remained the principal limit on measurement threshold at audio frequencies. The advent of low-noise, low-input-capacitance field-effect transistors (FET's), however, makes possible—for a given transducer impedance—an order-of-magnitude improvement at low audio frequencies, as compared with previous preamplifiers. These FET's also permit comparable performance at ultrasonic frequencies. This is shown in Fig. 1, the curves of which demonstrate that the equivalent input noise of the FET is far less sensitive to increasing transducer impedance than that of the bipolar (injection) transistor. These two properties of FET noise make practical the selection of a hydrophone of higher impedance—and therefore higher voltage sensitivity, other factors being equal—without an excessive raise in amplifier noise level. The result is a hydrophone-amplifier combination that is considerably improved over the whole usable spectrum, as far as measurement threshold is concerned. Reference 1 discusses the available tradeoffs in more detail. Figure 2 shows the equivalent noise pressure spectrum of one hydrophone-preamplifier combination. Compared to the best combination tried with a bipolar (injection) transistor preamplifier, it makes possible a self-noise reduction of at least 10 dB at audio frequencies, with some improvement at ultrasonic frequencies.



Bipolar Transistor Preamplifier



Field-Effect Transistor Preamplifier

FIG. 1. Equivalent Input Noise Spectra for Several Source Capacitances.

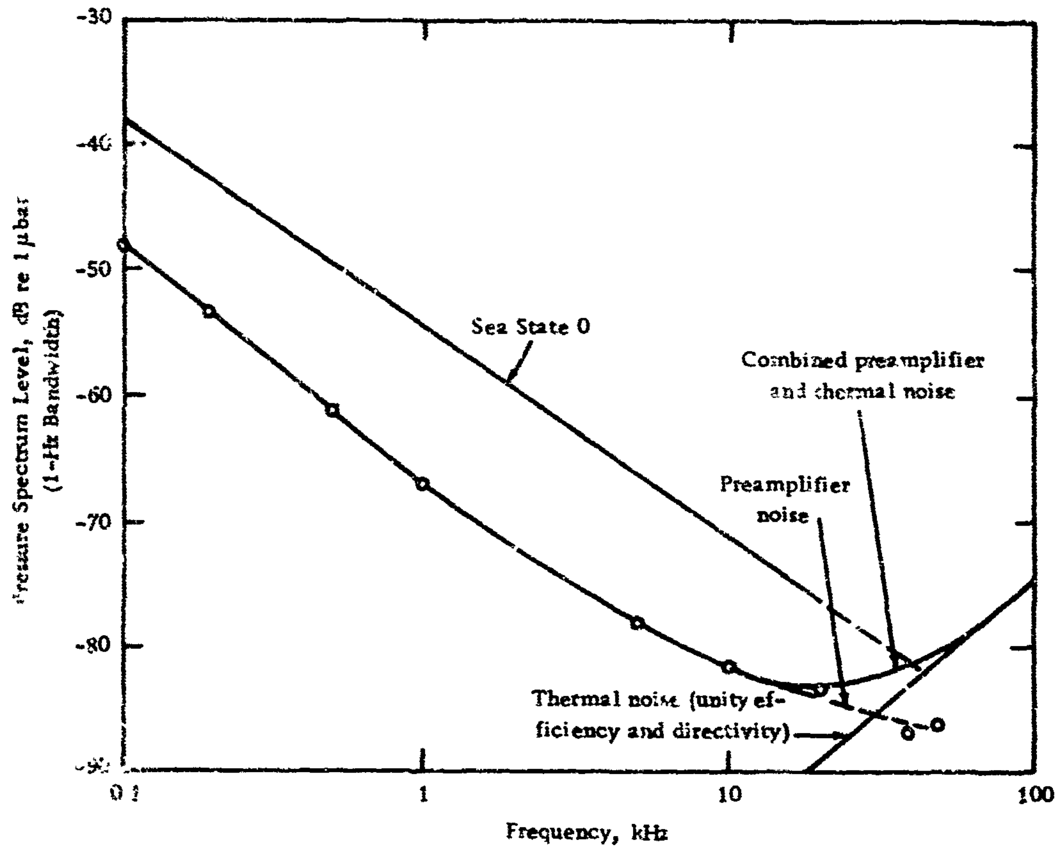


FIG. 2. Equivalent Noise Pressure Spectrum of Hydrophone Having Constant Sensitivity of −80 dBV per μbar and Capacitance of 60 pF.

CRITERIA FOR LOW-NOISE OPERATION OF THE FET

The measurement and characterization of FET noise is discussed in Ref. 2, 3, and 4. It should be remembered that, in the case of a ferroelectric transducer operating well below resonance, the generator impedance appears mainly capacitive. For this reason it will, in general, be impossible to optimize the generator impedance at more than one frequency for a particular amplifier. One approach, as demonstrated in Fig. 2, is to make the equivalent noise pressure spectrum of the hydrophone-preamplifier combination parallel to the slope of the lowest acoustic spectrum to be measured (in this case, the Knudsen Sea State 0 curve). At frequencies approaching transducer resonance, or at which the directivity factor is not unity, the

curve of Fig. 2 must be corrected. In the case of a small, wide-band hydrophone, however, the correction may not be appreciable below frequencies at which the thermal agitation of water becomes the controlling limit on measurement.

Laboratory measurements in advance of the circuit design confirmed the following criteria for low-noise operation of the FET in the frequency region where the device noise exhibits a 1/f relationship:

1. Other factors being equal, n-channel FET devices are inherently quieter than p-channel devices, because of greater carrier mobility
2. The DC bias point must be chosen to maximize the device transconductance; i. e., with V_{GS} approximately zero and I_D relatively high, so that operation is in the "pinch-off" or "drain-saturation" region (Fig. 3)
3. The gate-biasing resistor should be much larger than the transducer impedance at any frequency of interest, so that no significant dissipative element is added to the input circuit
4. Any un-bypassed resistance in the source lead must be kept small, as its effect is to add thermal noise in series with the input
5. The open-loop gain of the first stage must be large so that the noise contribution of the second stage is negligible

These criteria are satisfied by the circuit of Fig. 4a.

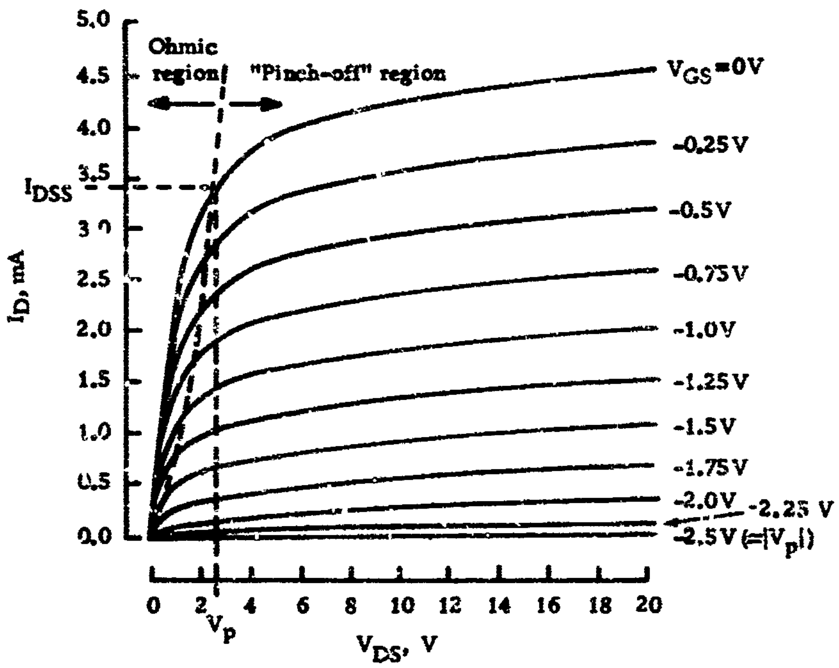
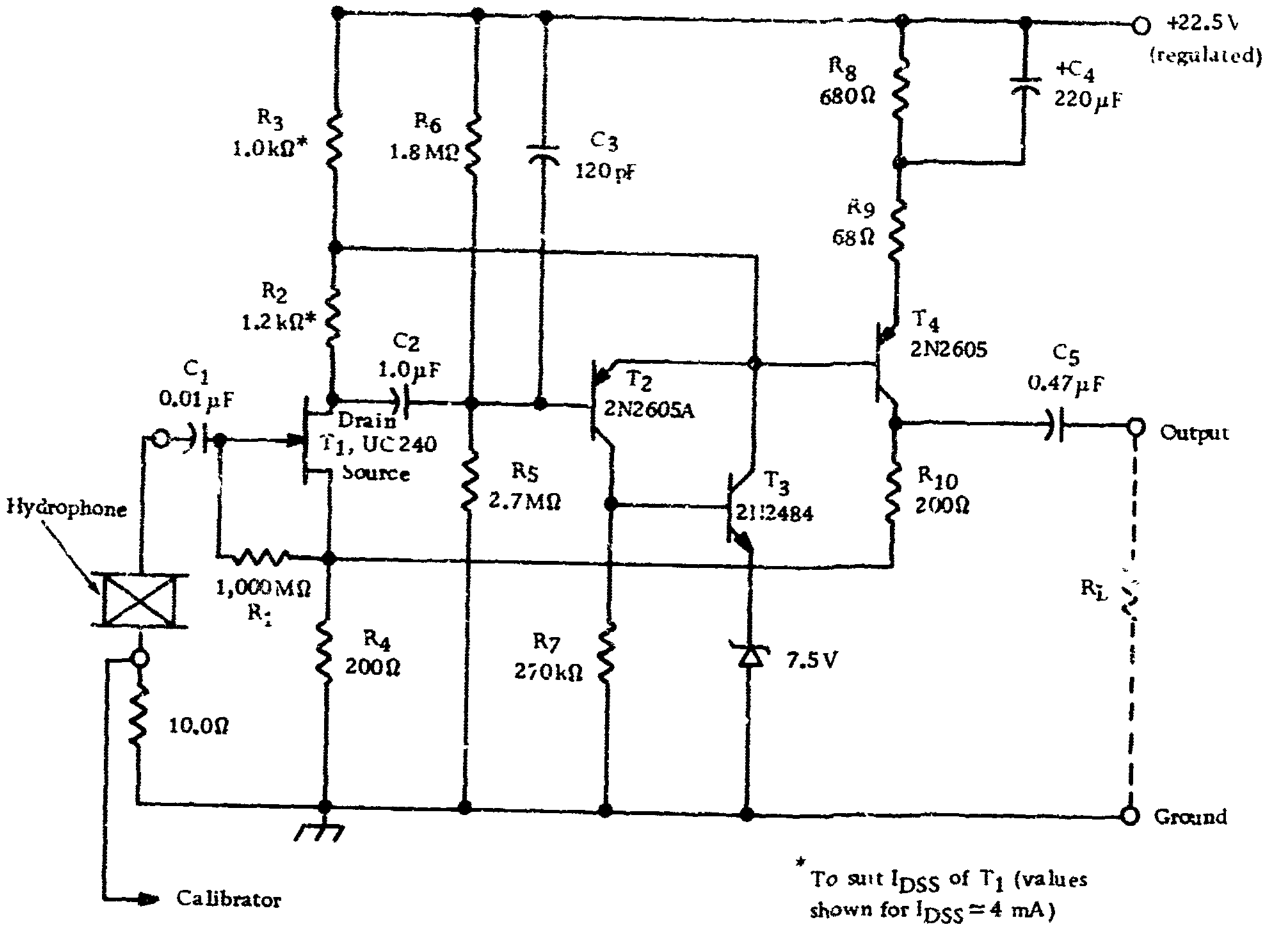


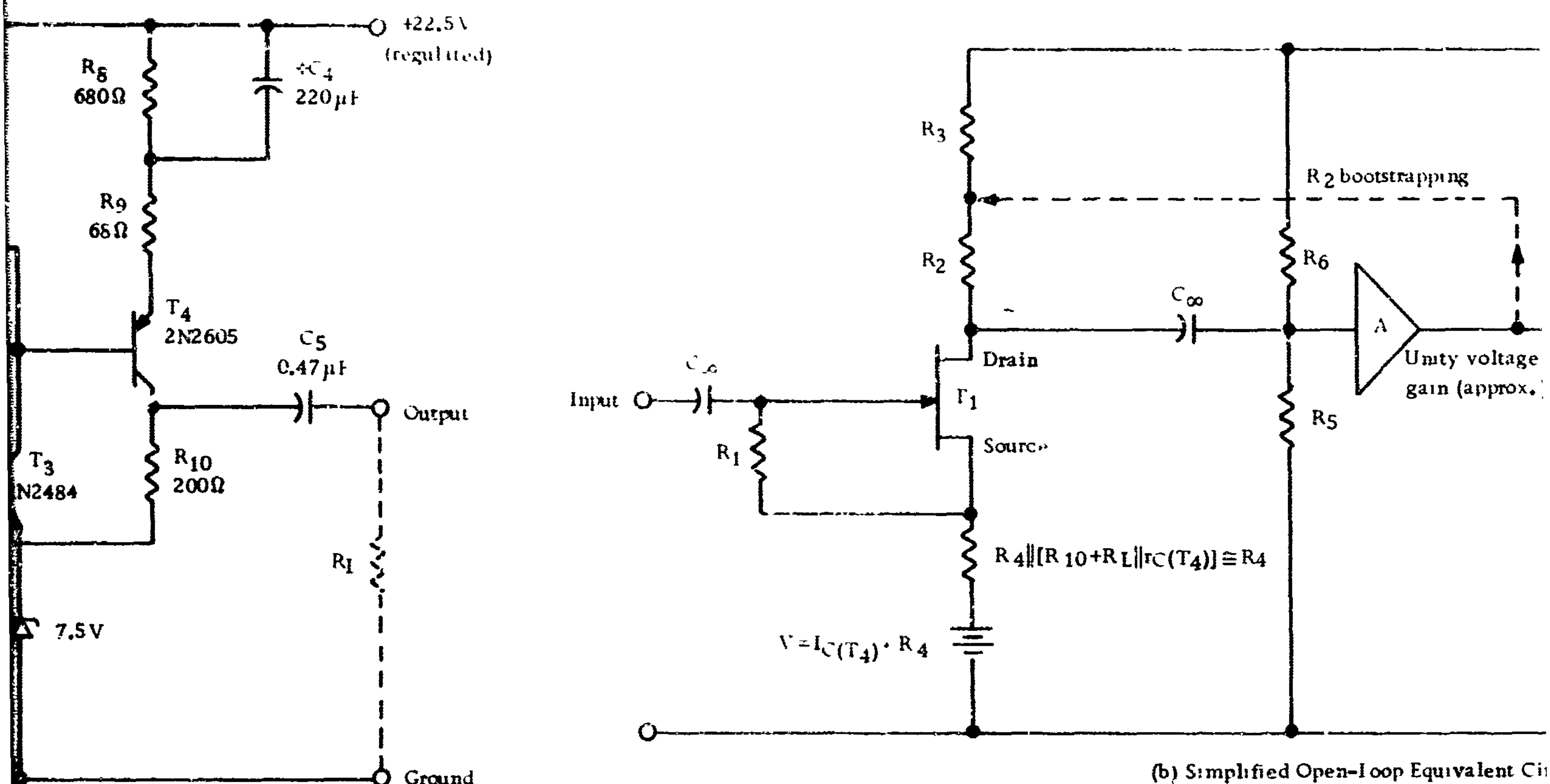
FIG. 3. Typical Drain Characteristics (Type UC 240 FET).



(a) Schematic.

FIG. 4. Pr

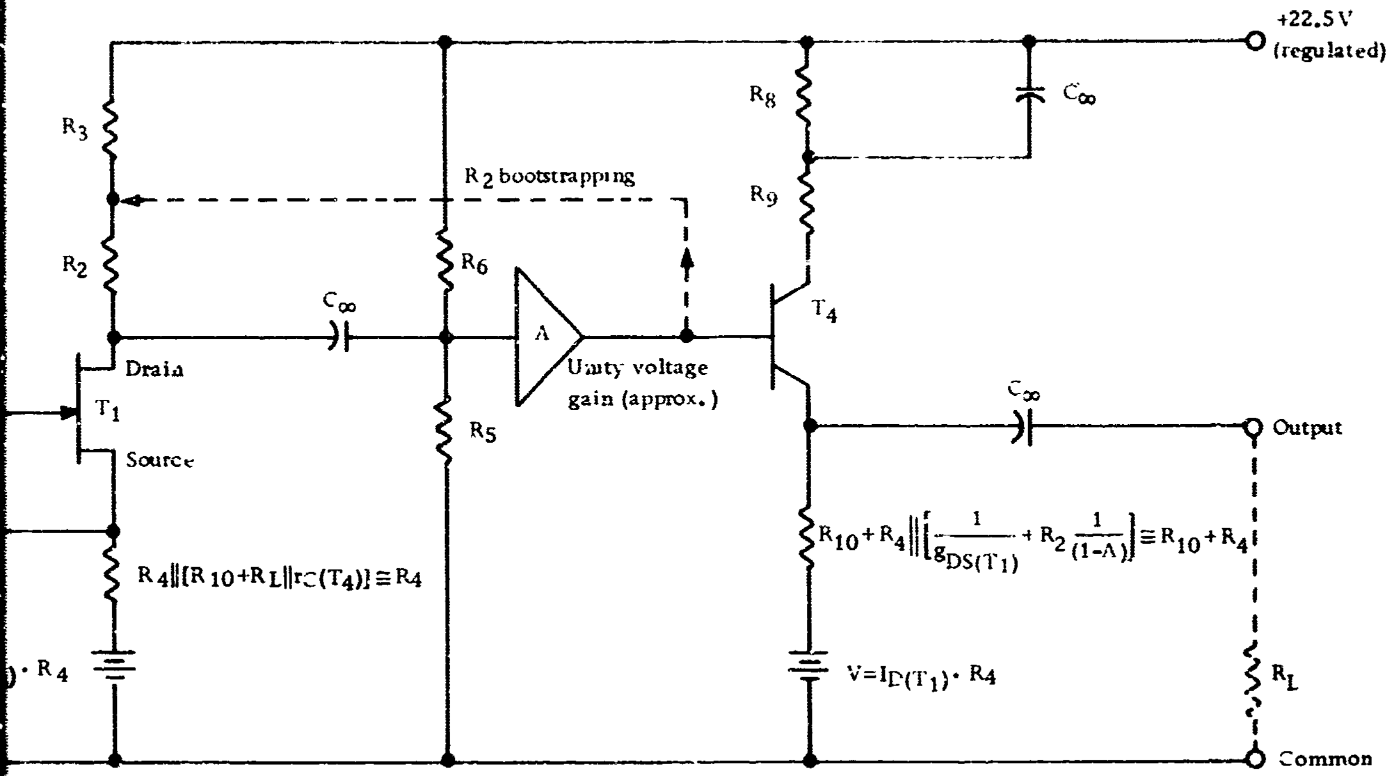
A



*To suit I_{DSS} of T_1 (values shown for $I_{DSS} \approx 4$ mA)

FIG. 4. Preamplifier Circuits.

B



(b) Simplified Open-Loop Equivalent Circuit.

its.

C

CIRCUIT DESIGN

The preamplifier circuit may be reduced to its basic concepts by opening the over-all feedback loop and considering the three parts shown in Fig. 4b: (1) the FET, T_1, operating in the common source mode; (2) a unity-gain voltage amplifier consisting of bipolar transistors T_2 and T_3; and (3) bipolar transistor T_4 operating in the common emitter mode. Closing the over-all loop sets the voltage gain at 6 dB by inverse feedback from the collector of T_4 to the source of T_1. Since negative feedback does not alter the inherent noise properties of an amplifier, it is convenient for purposes of understanding circuit operation to consider the amplifier in its equivalent open-loop condition. Furthermore, criteria established for low-noise operation of an amplifier without feedback also hold for an amplifier with feedback, provided that (1) the thermal noise and loading effects of the feedback network are taken into consideration; and (2) the feedback phase is correct (i. e., amplifier amplitude response is flat) over a band considerably greater than the desired signal bandwidth. If (2) is not fulfilled, the signal-to-noise ratio within the signal bandwidth can be seriously degraded. Gain peaks may appear at noise frequencies outside the desired band High-level noise outside the band will be only partial rejected because of the finite slopes at the edges of the passband. High-level noise outside the band can also appear within the band as the result of nonlinear effects (e. g., intermodulation) in the amplifier.

Besides the requirement for lowest possible noise, the following features were desired:

1. DC bias stability from 0 to 50°C
2. 6-dB voltage gain (feedback controlled)
3. Flat response from 10 Hz to 100 kHz
4. Input impedance 500 MΩ (minimum) at low frequencies

None of these presents a special problem when the circuit of Fig. 4a is employed.

The FET selected is the UC 240, an n-channel device having low-noise properties specified over a wide range of input circuit impedance (such as that encountered in broadband, omnidirectional hydrophones). The gate-biasing resistor (R_1) sets V_{GS} at very nearly 0 volt. A value of 1,000 MΩ was chosen for R_1, since the use of hydrophones having a capacitance as low as 15 pF was contemplated. This is a rather high value for such a resistor, but bias stability is easily attained, since operation is in the flat portion of the drain characteristic (Fig. 3), well above "pinch-off." In this region, [illegible] t, it is possible to operate successfully over a large temperature range without using an external gate-biasing resistor (Ref. 2). Total gate leakage current for the UC 240 is specified as 10^{-10} A (maximum) at 25°C and $V_{GS} = -30$ volts. However, since the gate-source junction has approximately zero voltage across it, and the gate-drain junction no more than 8 volts, it appears that about half the specified value of I_{GSS} (maximum) would be conservative. Thus, with $R_1 = 10^9$, V_{GS} could be as

great as +0.05 volt at 25° C or +0.2 volt at 50° C. The latter condition would probably not be encountered in a hydrophone, but even then the FET would function satisfactorily as a small-signal amplifier, since appreciable gate current is not drawn until the gate-source junction is forward-biased by about 0.3 volt (Ref. 3).

The drain current is free to change with temperature, but the change is less than ±10% over the range 0 to 50° C, and its effect on the circuit voltages is minimized by keeping the DC value of drain-load resistance (R_2) fairly small. Since operation in the "pinch-off" region is required, it is necessary to choose R_2 to suit the I_{DSS} value of each FET. Fortunately, the exact value of V_{DS} does not affect noise performance as long as it is greater than or equal to V_P (V_P = 5 volts maximum in this case), so that a particular value of R_2 will suffice for a range of at least 2-to-1 in I_{DSS}.

R_4 (200 ohms) does not enter significantly into the FET bias considerations. Its purpose is to provide a means of inserting negative feedback (source bootstrapping) when the over-all loop is closed. It is fairly small compared to the minimum equivalent noise resistance of the FET (in this case typically 1 kΩ), and therefore adds little noise to the input circuit.

Transistors T_2 and T_3 comprise a unity-gain amplifier that serves several purposes. First, it bootstraps R_2 so that the AC drain-load resistance for T_1 is high, allowing high open-loop gain for the first stage; at the same time, it establishes a DC voltage at the junction of R_2 and R_3, thereby stiffening the drain bias point of T_1. The amplifier also establishes the DC bias point of T_4, and provides power gain to couple the signal from the drain of T_1 to the base of T_4. The pair T_2-T_3, for practical purposes, may be thought of as a high-quality emitter-follower (i. e., one whose properties surpass those of the Darlington configuration). The circuit is described in detail in Ref. 5. T_2 is biased for low-noise operation ($I_C \simeq 30\mu A$), while T_3 is biased at a current comparable to I_{DSS} of T_1, so that it is capable of maintaining the DC voltage at the junction R_2-R_3 despite appreciable changes in I_D. If R_2 is raised to accommodate those FET's that have low values of I_{DSS}, R_3 should be increased by about the same ratio to prevent T_3 from having to pass more current than necessary.

Transistor T_4 is a conventional common-emitter amplifier, whose collector-load resistance is divided in two, with half of it (R_4) connected in series with the source of T_1 to provide a closed-loop voltage gain of 6 dB from the gate of T_1 to the collector of T_4.

AC stability at band-edge frequencies can be a problem with circuits of this complexity, especially when input or output terminations are reactive. The approach taken to achieve stability in this case was the common one of making one rolloff predominate at each end of the band. The predominant low-frequency rolloff is due to R_9 versus C_4, about 10 Hz. The predominant high-frequency rolloff is due appoximately to the dynamic drain resistance of T_1(dV_{DS}/dI_D for constant

V_{GS}, order of 30 kΩ) in parallel with the much higher bootstrapped value of R_2 versus C_3, about 60 kHz. In the application for which it was designed (Ref. 1), the preamplifier input is connected to a hydrophone, which is not only reactive but changes its reactance widely in the vicinity of resonance. With the indicated value of C_3, however, it has been possible to use a variety of hydrophones without introducing high-frequency instability. The output of the preamplifier is normally connected through RC band-limiting networks to a 20-dB balanced amplifier that drives the underwater cable. It is not recommended that this preamplifier be used to drive a cable directly, even though its output impedance is low, because the reactive loading could cause high-frequency instability.

PERFORMANCE

The main effort was spent on laboratory measurements to determine the most effective means of operating an FET as a low-noise amplifier for small hydrophones. This circuit was designed in accordance with the previously described approaches combined with simplifying approximations; no claim is made that it represents an "optimum configuration." It is the author's opinion that exact analyses of circuits of this complexity serve little purpose. Not only is the work time-consuming, but so many approximations have to be made to obtain tractable solutions that it is frequently impossible to decide whether a given solution is valid.

Modern computer techniques make it possible to perform linear analysis of circuits such as the one described, but care is required to see that the models used incorporate all the necessary parameters. It is also mandatory to make detailed laboratory measurements of each active device at its in-circuit bias point, since most parameter variations with bias conditions are not usually given in transistor data sheets. Furthermore, some knowledge of the parameters' statistical distribution among transistors of the same type is required, so that several sets of measurements must be made. When design time is limited, a pragmatic, experimental approach can be used.

In this case, a computer analysis is in progress, primarily for the purpose of testing the ability of the program to handle a circuit of this type, but also to check the adequacy of the transistor models.

Once it was determined that the FET could meet the noise requirements, it was necessary only to add circuitry to provide the feedback-controlled gain, flat response, and bias stability that are required of a practical amplifier and, at the same time, avoid degrading the noise performance of the FET. This effort succeeded, since the equivalent noise resistances measured with the complete preamplifier were at least as low as those measured in the laboratory test circuit.

As indicated earlier, it is necessary to preserve correct phase response over more than the required signal bandwidth to prevent

the negative feedback from degrading the noise performance of the FET. This means that the amplitude response will be far greater than that actually required, or desired, for the signal bandpass. Such is the case with this amplifier (Fig. 5). Simple resistance-capacitance (RC) networks ahead of the cable-driver amplifier ordinarily limit the signal bandwidth to about 25 Hz to 200 kHz (−3-dB points).

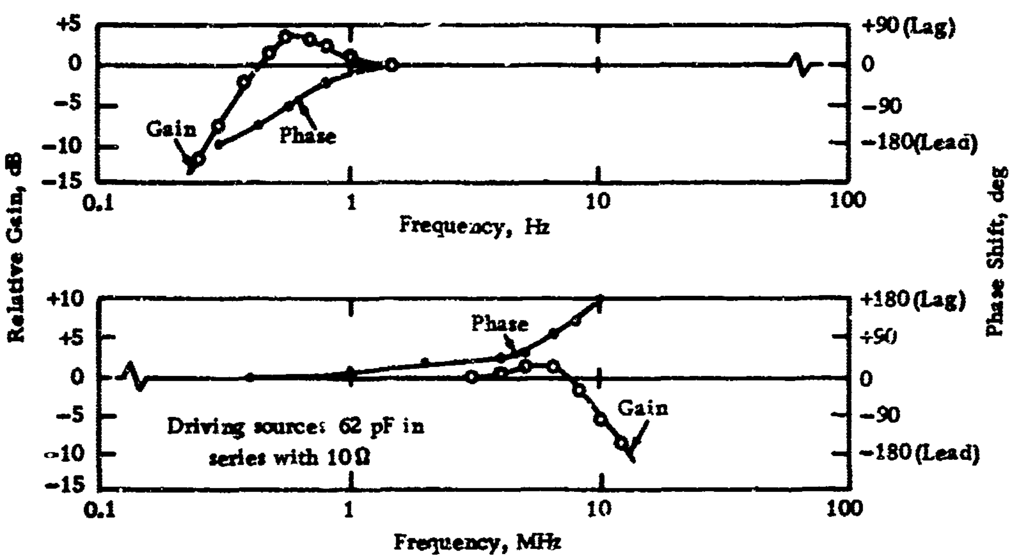


FIG. 5. Preamplifier Gain and Phase Response.

Typical transient response and overload recovery characteristics are shown in Fig. 6 and 7, respectively. The 10-volt peak-to-peak

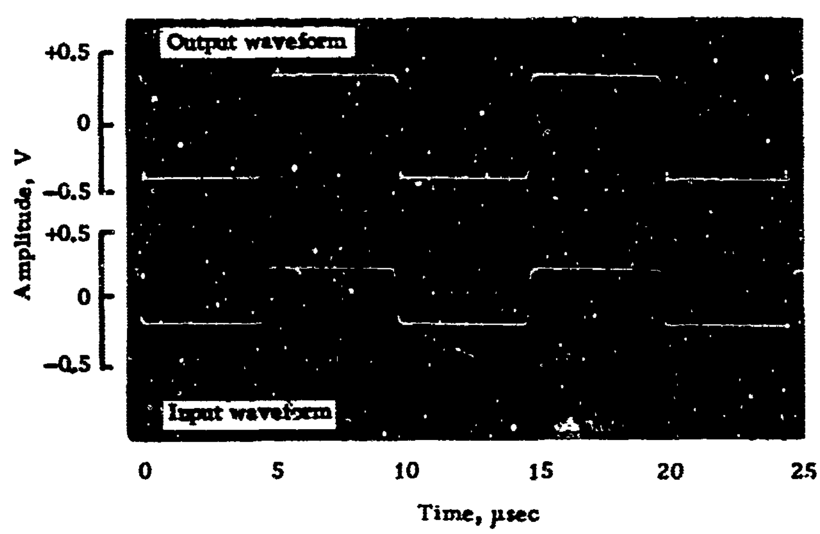
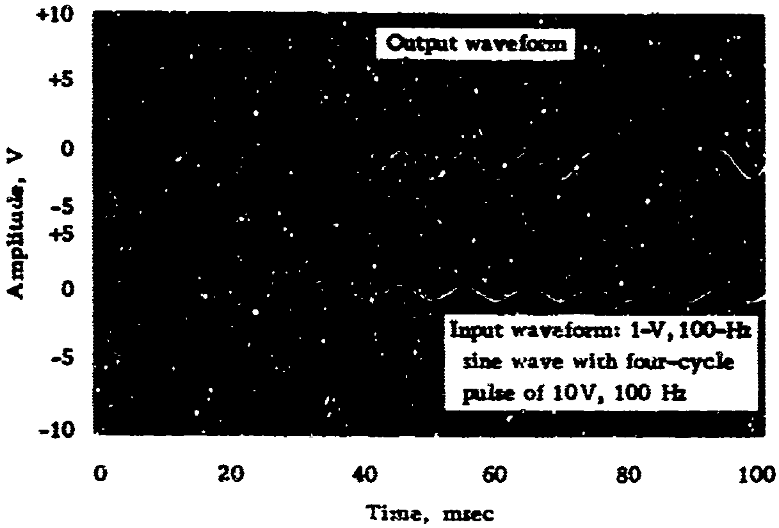
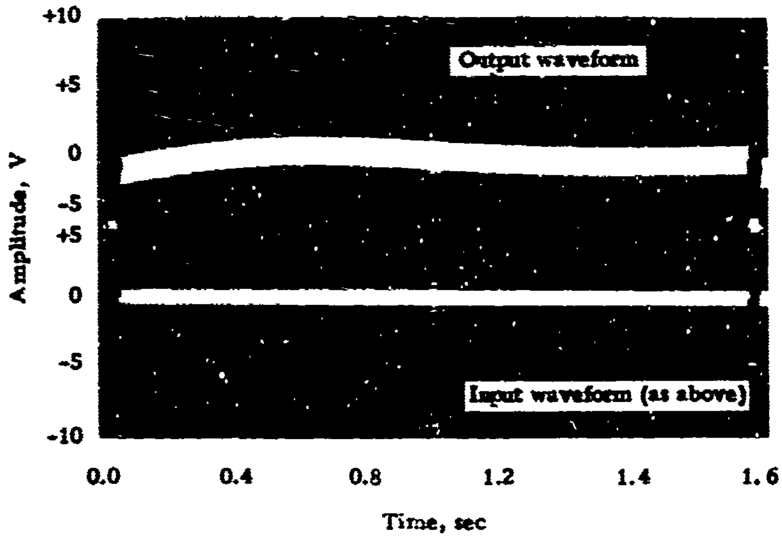


FIG. 6. 100-kHz Square-Wave Response.



(a) Waveform Distortion



(b) Baseline Distortion

FIG. 7. Overload Recovery.

input pulse in Fig. 7 represents nearly a 40-dB overload relative to the intended maximum signal level, as determined by saturation of the cable-driver amplifier.

The input resistance, measured at low frequencies (on the order of 100 Hz) is approximately 1,000 MΩ, and the input capacitance, measured at ultrasonic frequencies (on the order of 40 kHz), is about 5 pF.

CONCLUSIONS

The result of this project has been to produce a preamplifier yielding performance superior to those previously available. It has been used successfully in the field for nearly two years with various high-impedance hydrophones covering the band from about 10 Hz to 150 kHz. It makes practical a 100-kHz-bandwidth hydrophone-preamplifier system whose self noise is 10 dB below Sea State 0 ambient pressure from under 100 Hz to nearly the thermal noise limit. At the same time, it allows the use of a hydrophone small enough to preserve omnidirectionality up to approximately 30 kHz.

The other design objectives were met or exceeded. DC bias stability is maintained between 0°C and 50°C and voltage gain is 6 dB (±0.5 dB) from about 1 Hz to 1 MHz. Input impedance is approximately 1,000 MΩ at low audio frequencies, and input capacitance at ultrasonic frequencies is about 5 pF. Thus the preamplifier is suitable for use with very-low-capacitance (high-impedance) hydrophones.

Since the preamplifier noise is primarily dependent on the noise properties of the FET, the performance of this circuit could be improved by substituting improved FET devices as they become available.

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1. ORIGINATING ACTIVITY (Corporate author) U.S. Naval Undersea Warfare Center 3202 E. Foothill Blvd. Pasadena, Calif. 91107	2a. REPORT SECURITY CLASSIFICATION UNCLASSIFIED 2b. GROUP

3. REPORT TITLE

A LOW-NOISE FET PREAMPLIFIER FOR WIDE-BAND HYDROPHONES

4. DESCRIPTIVE NOTES (Type of report and inclusive dates)

Development report

5. AUTHOR(S) (Last name, first name, initial)

Allman, Richard I.

6. REPORT DATE	7a. TOTAL NO. OF PAGES	7b. NO. OF REFS
March 1968	12	5

8a. CONTRACT OR GRANT NO. Bureau of Naval Weapons Task Assignment RUTO- b. PROJECT NO. 3E-000/2161/F108-03-08 (Problem No. 115) c. d.	9a. ORIGINATOR'S REPORT NUMBER(S) NUWC TP 24 9b. OTHER REPORT NO(S) (Any other numbers that may be assigned this report)

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13. ABSTRACT

The design of a low-noise, field-effect transistor (FET) preamplifier is described, and criteria are given for low-noise operation of relevant FET devices. When used in conjunction with a hydrophone having a sensitivity of −80 dBV/μbar and a capacitance of 60 pF, the design provides a measuring system having a self-noise threshold approximately 10 dB below Sea State 0 from less than 100 Hz to nearly the thermal limit. At the same time, DC bias stability is maintained between 0°C and 50°C, and voltage gain is 6 dB with frequency response essentially flat from 1 Hz to 1 MHz. Input impedance is approximately 1,000 MΩ at low frequencies, and input capacitance at ultrasonic frequencies is approximately 5 pF.

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